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A D.C. amplifier with small drift and adjustable gain.

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A D. C. AMPLIFIER WITH SMALL DRIFT

AND ADJUSTABLE GAIN

by

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TABLE OF CONTENTS

ABSTRACT

Title of Thesis: A D. C. AMPLIFIER WITH SMALL DRIFT AND ADJUSTABLE GAIN

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The problem of amplifying slowly varying signals is an old one. The use of vacuum tube amplifiers for the objective is considerably complicated by the fact that the low frequency of the signals precludes the use of capacitive coupling between the stages and this brings forth the problem of drifting.

Various solutions have been proposed, each one with a definite application.

One obvious way of making possible the use of the common R-C coupled amplifier is to modulate the signal we want to amplify and at the end to demodulate it through a phase sensitive demodulator.

This thesis is concerned with such an application, using a 400 cps "chopper" as modulating device and a vacuum tube switch as a phase sensitive full wave demodulator. Feedback is used in conjunction with a high gain AC coupled amplifier to stabilize the gain. The device has a low output impedance, of the order of 10K, and a high input impedance, of the order of 1M.

Other specifications are:

a) Three separate, additive inputs.

b) Adjustable gain in four steps, namely $k = 1, 2, 5, 10$.

c) Gain stability better than 98%, namely:

$$\left| \frac{\frac{e_0}{k} - (e_1 + e_2 + e_3)}{e_1 + e_2 + e_3} \right| \leq 0.02 \text{ for frequencies between 0 and 90 cps.}$$

Good results were obtained.

HISTORY OF THE PROBLEM

Troubles with d. c. amplifiers are too widely known to deserve an extensive comment.

It is enough to say that the tendency to drift and generally varying conditions of operation of such devices makes them unreliable to such type of application in which no constant adjustment is allowed.

Besides, even in the situations they have to be used they still are not ideal devices due to strict conditions they impose on the power supply, to their price and to the continuous attention they require.

One obvious means of improving the drift problem is to use an R-C coupled amplifier. Since the required size of the components makes this type of amplifier unsuited for low frequency amplification, the solution it to modulate the signal, amplify the modulated signal and demodulate it after the amplification.

Several methods can be chosen for the modulation and corresponding demodulation.

When the frequencies to be amplified are not too high switching modulation using a mechanical type contactor presents some advantages. A significant gain in stability is achieved as compared with circuits using vacuum tubes - the modulating circuit is very simple, the contactor consumes very little power and doesn't need too much space. Since nearly all contactors are single-pole, double-throw, they are suitable for full wave modulation.

The Rev. Sci. Instr. (17, 194 - 1946) has a report from Liston et al describing an amplifier to be used as a substitute for a very low

resistance galvanometer, having two synchronized contactors for modulation and demodulation.

Approximately on the same basis, the A. I. E. E. Transactions (67, Vol. I, P. 47) brings a report from Williams, Tarpley and Clark, describing a d-c amplifier stabilized for zero and gain, to be used for measurements of very small voltages and currents. Variations of the basic design for low and high impedance input circuit and for series and parallel feedback are briefly described.

Several books make references to the possibility of using contactor modulation to improve the qualities of stability in d-c amplifiers.

Theory of Servomechanisms (James, Nichols and Phillips) dedicates a few pages to contactors and their applications.

Waveforms (Chance, etc.) describes two types of mechanical contactors and points out their advantages over their electronic counterparts.

No more literature was found treating of this particular problem.

SUMMARY

Servo systems have a need of amplifying devices for electrical signals which have a good response up to about 40 cps and which can be considered to have as small interaction as possible with the neighboring elements in the system, namely high input inpedance and low output impedance.

We have proposed to design an amplifier which will follow as much as possible the following characteristics.

1- Three separate additive inputs, besides the feedback

2- Input resistance of at least 1 M for each input

3- Adjustable gain in 4 steps: 1, 2, 5, 10

4- Accuracy better than 98%, namely:

$$\left| \frac{\frac{e_0}{K} - (e_1 + e_2 + e_3)}{e_1 + e_2 + e_3} \right| \leq 0.02$$

for sinusoidal inputs of frequencies from 0 to 40 cps

5- Output resistance of 10^4 ohms or less

6- Stabilized power supply is not provided

In starting the planning of such an amplifier the paper of Williams et al* was thoroughly read, since his problem has points of similarity with ours.

The main differences are:

a) William's amplifier is designed for exceedingly low level signals;

b) It is supposed to amplify essentially D. C. signals.

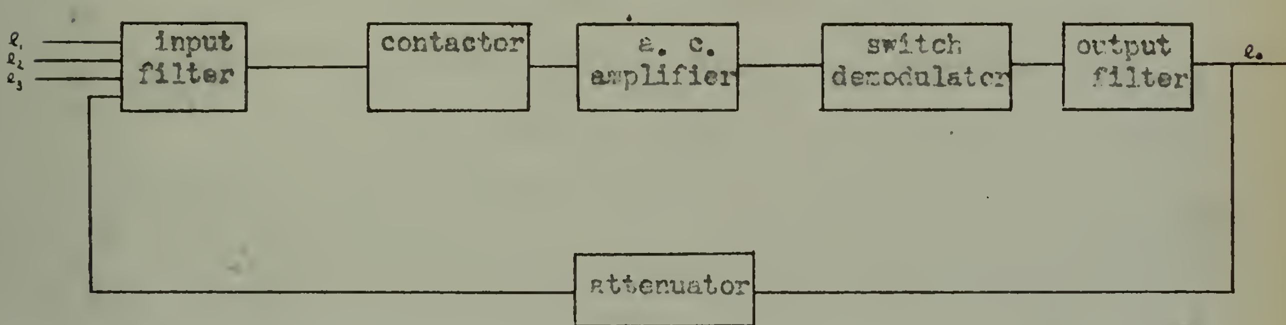
* Reference 1

c) The power level at the output is sufficiently low to permit the use of a chopper to demodulate the signals.

With those ideas in mind we tried to modify William's project to our objective and simplifying it as much as possible to eliminate the most of the essentially "precise" features, such as shielding the amplifier, stabilizing the power supply, using high accuracy components, etc.

Description of the system

A block diagram of the system follows and after that the components will be commented upon one after another.



Input filter - It is supposed to attenuate all the frequencies out of the chosen range, especially 400 cps since a 400 cps signal coming in will go through and have the same effect as a D. C. input.

At the same time it must provide for the input impedance we wish.

Contactor - A 400 cps Stevens Arnold mechanical "chopper" was chosen. The frequency of 400 cps has two good advantages: a) it is high enough to use an a. c. amplifier to good advantage even considering

the bandwidth of 80 cps; b) Its ratio to the maximum frequency we want to amplify is high enough to allow good amplification.

The "chopper" itself didn't show ideal qualities for the low level input we wanted to use. It presented a slight zero error - besides that it seemed very stable in what refers to the duty cycle, namely the fraction of a cycle. The contact remains open or closed.

A.C. amplifier - A 4 stage amplifier using a 6L6 as output was designed.

Two 6SL7's twin triodes were used, the socket of the first one was mounted on rubber to decrease mechanical pick-up.

The first tube was connected as a cathode follower and this connection proved a big improvement over the regular grid connection. Although the reason wasn't completely understood it seemed to be related to disturbances on grid current as caused by the switching action.

The circuit was designed to have optimum transmission at 400 cps and low drain when no signal was applied.

A potentiometer was used between the two 6SL7's in order to vary the gain of the system at will.

In the output to the demodulator a step-up transformer was used to match impedances.

Switch demodulator - broadly speaking it consists of 2 separate parts: the synchronizing circuit and the demodulator itself.

The synchronizing circuit uses the other side of the s.p.d.t. chopper to provide a signal which is further amplified by a 6SN7 and goes to the demodulator across a step up transformer. A capacitor is used

across the primary to provide adequate phase shift for correct synchronization.

One advantage of this synchronizing circuit over the direct connection of the 400 cps carrier to the demodulator is that any phase shift due to mechanical shock or otherwise which occurs at the mechanical switch will not affect the synchronization since it will shift the synchronizing signal at the same time. The demodulator is a four double diodes circuit described in page 402 of Waveforms* and which is the electronic equivalent of the "chopper". Its advantage over the "chopper" is the fact of greater power capability. Its disadvantage is the non linearity, which will cause the common distortion characteristic of diode rectification.

Output filter - only a capacitor was used across the load resistor, to filter the 800 cps ripple from the rectification.

A value as small as possible was chosen in order not to produce undue attenuation to the signal.

Attenuator - The overall gain is chosen through this attenuator.

In the experimental model it was just a potentiometer but in a final set-up it should consist of 5 resistors in series; their values should be chosen in such way that when the feedback is connected between 2 of them through a multiple throw switch, the amplifier will have the correct gain.

Theory of Operation

Although our research was almost entirely of experimental nature some theory had definitely to be used in order to see whether or not the procedure we were following was logical.

Some statements which are being and will be made in this Summary

* Reference 2

will be explained in more detail later on in this report.

The final a. c. amplifier used in the system was designed to have a calibratible gain, with a maximum of about 10^6 from the 1st grid to the secondary of the output transformer, across the load - making use of the simplifying principles laid out in Applied Electronics * and with data from the RCA tube manual, the amplifier was designed with a flat range of amplification (within 5%) from $f_1 = 290$ cps to about $f_2 = 60K$. This large upper frequency comes essentially from the fact that triodes with a very low inter electrode capacitances have been used.

Since we can definitely correlate gain attenuation with phase shift those data will indicate a very small phase shift above and below 0° for frequencies between $10f_1$ and $0.1\ f_2$ or 290 cps to 6500 cps, more precisely $5^\circ.8$ per stage at those limits.

This would give a phase shift of about 30° plus the phase shift imposed by the output transformer itself.

As a broad statement we can say that the a. c. amplifier will not create any problem of instability.

Moreover, for frequencies between 360 and 440 cps this phase shift will obviously be much smaller.

A Fourrier analysis of the input waveform shows that the "chopper" will create a modulated square wave which contains the sum and difference frequencies of the signal with each harmonic of the "chopping" frequency, plus an average voltage wave.

We will be concerned essentially with the fundamental or its side

* Reference 3

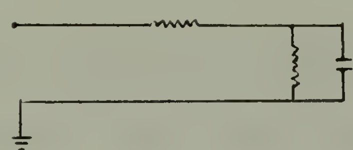
bands, since the average and the higher harmonics will be attenuated at the output transformer, demodulator and output filter.

They certainly will not have a marked effect on the gain since this harmonic effect deals almost equally with all the modulating frequencies we care for, which constitute a relatively small fraction of the carrier frequency.

If we are allowed to make an image, we would say that the range of frequencies we deal with are almost bunched at a point on the curve of transmission versus frequency, of the a. c. amplifier.

Having made some simplifications in the treatment of the signal loop, across the a. c. amplifier we still have to deal with the demodulator and output circuit, and input circuit in order to verify the possible phase shifts to the signal.

The demodulator itself is a purely resistive device. However a capacitor is used in paralled with the load resistance in such way that a simplified circuit for the signal from the secondary of the output transformer is as follows.



This condenser is used to smooth out the 800 cps ripple which comes from the full wave rectification.

At the worse conditions it would bring a 90^{0} phase shift which added to the amplifier phase shift would still be smaller than 180^{0}.

In the input circuit up to the "chopper" there is no capacitor and so we are fairly safe in the matter of stability. This statement is

made regarding as valid the simplifying assumptions made previously, namely, that the loop attenuation of the 400 cps "chopper frequency" and its harmonics is high enough to be neglected and that we deal only with low frequency signals which enter by no other place besides the input.

Under those ideal conditions the gain could be stabilized as much as we wished through feedback.

Thermal noise would be the limiting factor for maximum stable gain.

Actually the circuit behavior departed somewhat from those simplifying assumptions, due mainly to the 60 cps pick-up which could not be curbed entirely with the resources we could count with in the laboratory.

All the means we tried to use to decrease this effect will be described later.

The maximum loop gain obtainable for a pure D. C. signal was of the order of 40, and slightly less for A. C. signals.

The conclusion reached at the end of the investigation is that the system as planned is good and should be able to do the job it was designed to.

Improvements in the construction technique, such as good shielding, good ground and a decreasing of "zero error" will make this system very well suited for high gain and excellent stability.

DESCRIPTION OF THE PROCEDURE FOLLOWED IN THIS PROBLEM

In the description which follows no attempt will be made to describe all successes and failures nor to show how the time was spent in the project. Only significant results will be described.

At the start of the investigation, several questions were bound to arise.

How would a system like this behave in closed loop?

What order of magnitude would be possible to use for gain?

How carefully built should the components be?

Should a narrow band or wide band amplifier be used?

What should be used as a demodulator?

How difficult is the synchronizing problem?

What relative importance would have thermal noise, thermoelectric effects, 60 and 400 cps pick-up, magnetic induction, etc.?

To get a rough idea of how, the questions would be answered a system basically similar to the one originally planned was set up.

The main components of the system were:

a) Brown single pole double throw contactor excited by 18 volts, 400 cps;

b) Ballantine wide band amplifier;

c) Feedback Control Laboratory model E E S - 6 gated demodulator, using triodes.

Besides those components, the input and output circuit were the same as the ones to be used in the future, and can be found in the diagram.

The following voltage supplies were needed:

110V, 60 cps, for the Ballantine amplifier

250 V, d. c., positive and negative; 6.3 V, 60 cps; 110 V, 400 cps, all for the demodulator

18 V, 400 cps, for the "chopper"

It would be tiresome for the reader, the description of all our successes and failures before final conclusions were reached. The main points, however, will be commented upon in the same way as the final conclusions.

To start with, the over-all gain of the system seemed to be somewhat low. The Ballantine amplifier uses feedback and has an over-all gain of about 300. The demodulator also has some gain since it is a triode type demodulator which detects and amplifies at the same time.

The input circuit used, the low impedance output circuit plus the output filtering brought the over-all gain all the way down.

The figures obtained are as follows:

a) Output across a capacitor only.

Over-all gain for a d-c signal was about 450 with less than 2% ripple.

b) Output across a 25K potentiometer and a 2 mf capacitor. Over-all gain for a d-c signal was about 100 and for 20 cps signal was about 10.

It is evident that such a large difference in gain could not be compensated through feedback.

There was no stability problem with this circuit. The feedback ratio was varied at will.

The chief troubles met in this part of the work were:

a) "Chopper" - The operation of the "chopper" was erratic and sometimes several hours were spent trying to adjust its contacts. As soon as it would start to work the contacts would come out of adjustment again.

Besides, the 400 cps pick-up was somewhat higher than expected.

b) Demodulation - The synchronization of the reference and signal voltages was very difficult if not impossible to obtain. A R-C network was used in the 110V, 400 cps, synchronizing line, and adjusted to give the smallest ripple. Even so the results were very poor.

The zero output error could not be eliminated from the demodulator. By zero output error, we mean the output when the input to the demodulator is shorted.

A compromise calibration had to be used to give minimum possible ripple with minimum possible output error.

c) Other causes - Nothing else appeared to influence the operation of the system appreciably. 60 cps pick-up and other minor effects didn't show up to any great extent.

After those conclusions had been reached a new system having a higher gain was built, using now a new type of "chopper".

This new system was actually identical to the ultimate system appearing in the diagram, except for the first tube which was not present.

Some difficulty was experienced in getting hold of some of the desired components such as transformers and condensers. The ultimate choice was a compromise, namely, the choice fell on the best possible

components which could do the job and at the same time could be obtained at the Feedback Laboratory.

Following the practice of going from the simpler to the more complex no particular care was taken at the start for decoupling the power supply or preventing some other possible disturbances.

The a. c. amplifier worked very well by itself, except for the 60 cycles "hum"; even that was not too disturbing.

After the amplifier was checked, the 400 cps was connected to the "chopper" and a d-c signal was applied at the input.

The effect of the switching operation of the input of the first grid was greater than suspected, giving a peak voltage several times the amplitude of the square wave signal.

It was suspected that the synchronizing circuit was drawing excessive current and at the same being coupled to the signal circuit through the "chopper". Change of the load resistance of the synchronizing circuit square wave generator from 0.5M to 5 M improved the situation but didn't eliminate the trouble. A small 1400 $\mu\mu f$ capacitor between grid and cathode was the solution for the problem and this particular disturbance practically disappeared. As expected, this capacitor brought with it considerable attenuation for the input signal but it was a price worth paying.

The next trouble to be met was the matter of synchronizing the reference and the signal voltages at the demodulator. It was solved by a convenient choice of capacitors to be connected shunting the primary of both output transformers. Those condensers also accomplished the

simultaneous job of improving the 400 cps response of both circuits, by rounding the corners of the still-very-nearly-square waves at that point, and so preventing them from being differentiated across the transformers.

The following problem to be met was the filtering of the 800 cps ripple present at the output.

It was decided to use one condenser in parallel with the load resistance.

Its value was a compromise between an optimum high frequency attenuation and a small low frequency attenuation.

Finally the question of 60 cycle "hum" was coped with. It was decided that decoupling on the power supply was necessary.

This decoupling was performed and a good improvement was noticed.

After all those corrective operations were done the system was tested under actual working conditions.

At zero signal input there was d-c and 60 cps present at the output. However the amplification for both d-c and a-c up to 40 cps taken from an audio amplifier was of good quality.

The chief drawback of this system was the small over-all gain, due to the several attenuations scattered along the signal path.

The maximum possible gain was of the order of 25 and there was no trouble in increasing the feedback ratio to a point where the over-all gain was 1. No oscillations occurred due to such small loop gain used.

The final conclusion reached at for this particular system was that it was in general good. The maximum gain was however, small to insure

the gain stability through use of negative feedback the way we want to be.

This conclusion led to the decision of building another amplifier (which would be the final one) with somewhat larger amplification.

The availability of 6SL7's added to the fact that they have very good characteristics of gain and low interelectrode capacitances led to the choice of one of them to give the additional amplification. Since only one stage of amplification was needed the remaining triode was to be used as a cathode follower. This connection has the advantage of making the input capacitance of the tube much lower * than it is for the regular amplifier input connection. This lowered capacitance should decrease the input signal attenuation and, as would be noticed on the actual operation, helps to eliminate the transient effects of the switching operation.

An additional feature was the introduction of a potentiometer between the two 6SL7's to provide a variable gain to the amplifier.

The plate supply for all tubes was decoupled by using R-C networks.

The circuits for the a.c. amplifier and for the synchronizing circuits were calculated by our own judgment and with help of the RCA handbook.**

The a. c. amplifier was tested at first for static conditions of bias and supply voltages. A 60 cycle "hum" was present all along and due to the high gain of the amplifier it was enough to saturate the output.

* Reference 4

** Reference 5

The first place where it appeared was across the grid resistance of the input tube when nothing else was connected to the grid.

It was then decided to use rectified heater supply to the 6SL7's and this was accomplished by a bridge circuit using selenium rectifiers, plus suitable filtering. Grounding of one side practically eliminated the 60 cps trouble on the two initial stages. The size of the decoupling capacitors at the later stages was increased and more improvement was noticed.

Even so, the disturbance at the output tube was still quite troublesome and this was solved by grounding of the 6.3 ac. heater supply, across a 30 ohms resistor from each side of the line.

These a. c. disturbances coming primarily from the 60 cps supply couldn't be entirely eliminated until the end. They were decreased to the minimum value possible, without use of shields or any other elaborated construction method.

Another trouble which persisted until the end was lack of a good "ground". When we are trying to achieve a good operation at low level, any small variation counts.

As we can see it, the two facts above were the principal causes which limited the maximum obtainable loop gain.

After the a. c. amplifier was tested for a. c. operation, the 400 cps were connected to the "chopper" and a d. c. signal was applied to the input.

Good synchronization was verified at the output by observing the waveform across the unfiltered load with a scope.

The zero error of the "chopper" was somewhat decreased by grounding the center tap of a potentiometer connected across the 6 volts, 400 cps, supply.

Operation with d-c and with a-c up to 40 cps was tried before any feedback was attempted.

The additive characteristic of the device was tried by applying signals at the different inputs, measuring the output and verifying the linearity of the results.

At that point the feedback connection was made and the gain was measured. The maximum loop gain was still less than what we desired it to be, but no improvement was in sight right away and time was running short.

The gain can be calibrated from 1 to 10, as desired, as long as the amplifier gain is decreased low enough to prevent oscillation.

It can be increased to high values, but always bearing in mind that the feedback has to be decreased accordingly to conform itself with the maximum possible loop gain, which ultimately determines the stability, as we can recall from the principles of Nyquist.

DETAILED DESCRIPTION OF THE FINAL SYSTEM

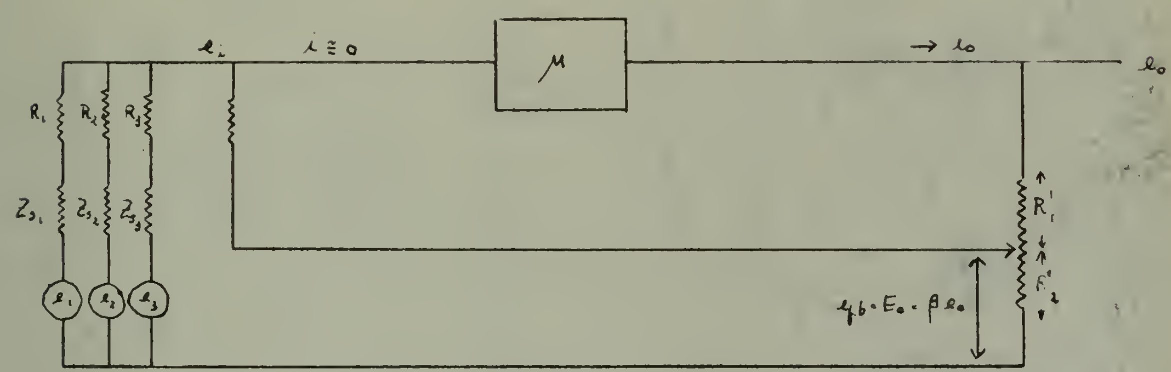
A. Diagrams

The two diagrams describing the final system as it was built are presented in the beginning to give the reader a referance for future reading.

They are presented in the next two pages.

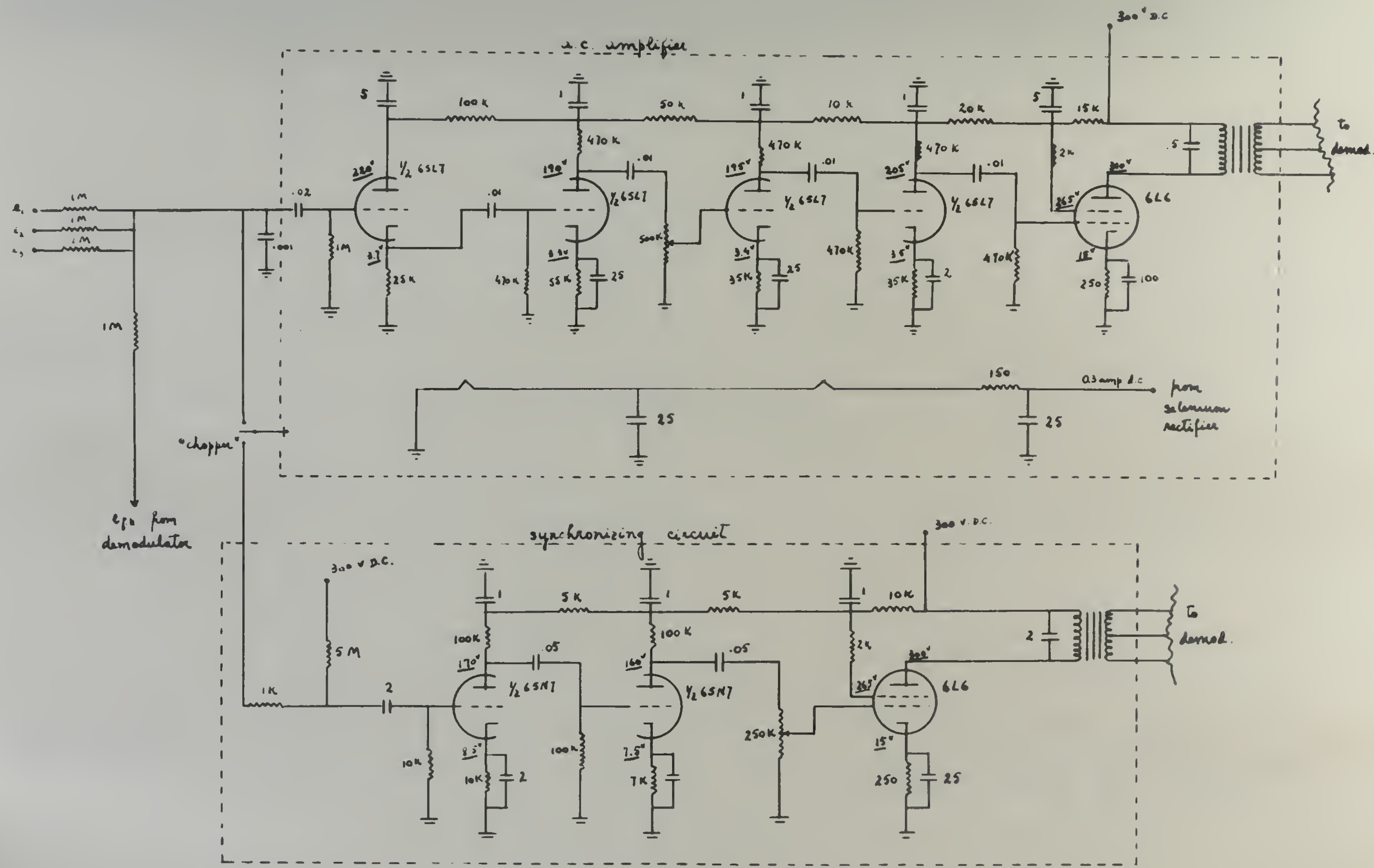
B. Analysis of the gain, output, impedance and input impedance

The following diagram is a good approximation of the circuit behavior and will be used for the analysis which follows.

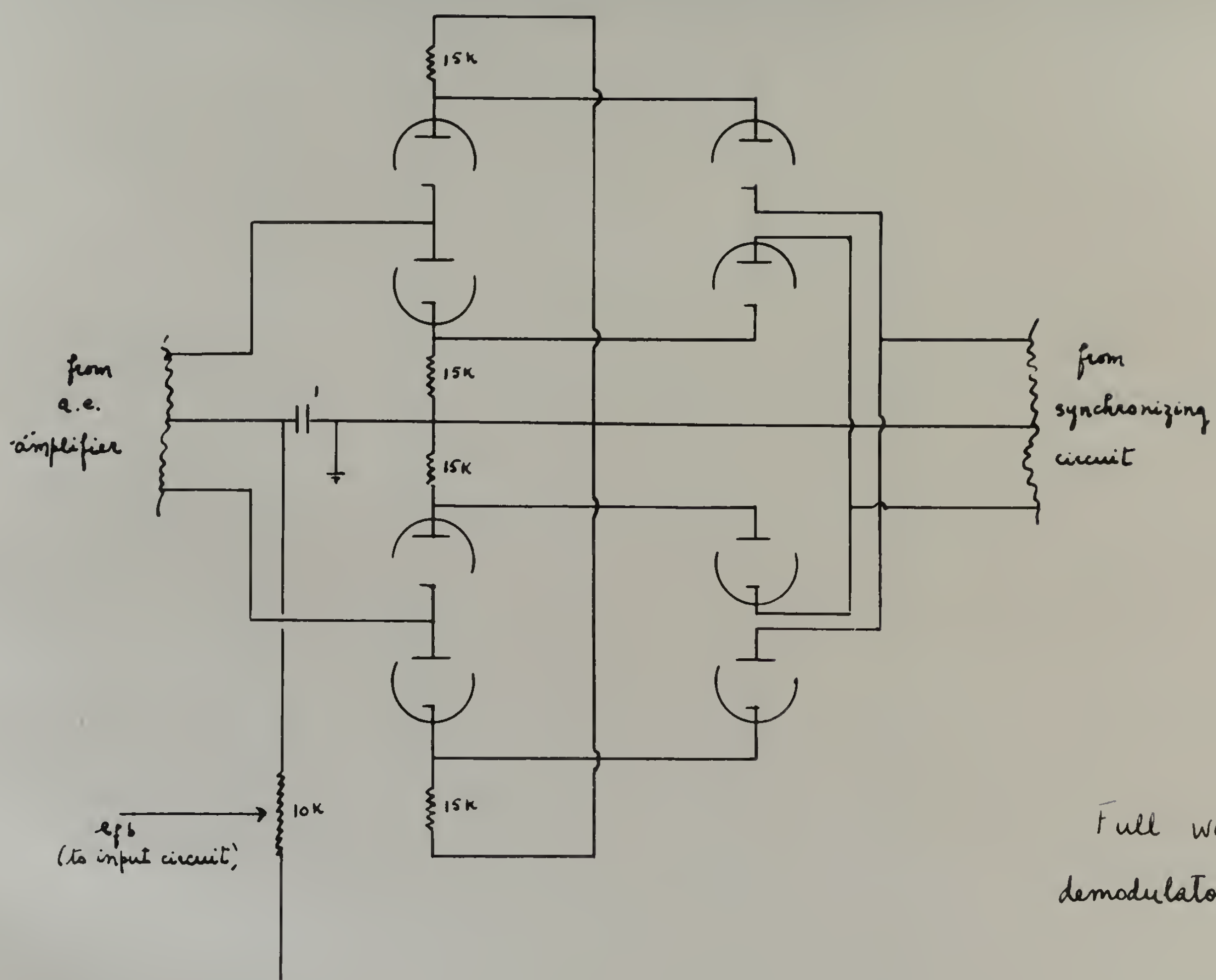


1. Gain

Due to the high gain of the amplifier, the input voltage will be, by necessity, very small. In this way the input current can be neglected in this analysis, since the input impedance for the a. c. amplifier is very high.



all capacitances in μf



Full wave, phase sensitive
demodulator, using four 6H6's

The equation for the input node can then be written:

$$\frac{e_1}{Z_{s_1} + R_1} + \frac{e_2}{Z_{s_2} + R_2} + \frac{e_3}{Z_{s_3} + R_3} + \frac{E_0}{R_4} = e_1\left(\frac{1}{Z_{s_1} + R_1} + \frac{1}{Z_{s_2} + R_2} + \frac{1}{Z_{s_3} + R_3} + \frac{1}{R_4}\right)$$

If $R_1 = R_2 = R_3 = R_4 = R$ and $Z_{s_1}, Z_{s_2}, Z_{s_3} << R$ we can simplify:

$$\frac{e_1 + e_2 + e_3 + E_0}{R} = \frac{4e_1}{R} \quad \text{or} \quad e_1 = \frac{e_1 + e_2 + e_3 + E_0}{4}$$

The above equation shows how the input signals add to from the input voltage to the amplifier.

We know that $e_0 = -\mu e_1$ where μ is the complex gain of the system and the minus sign is placed arbitrarily.

Then we may write:

$$-\frac{e_0}{\mu} = \frac{e_1 + e_2 + e_3 + \beta e_0}{4} \quad \text{or}$$

$$e_1 + e_2 + e_3 = \Sigma_e = \left(-\frac{4}{\mu} - \beta\right) e_0 = -e_0 \frac{4 + \mu\beta}{\mu}$$

Now the gain of the system can be obtained as:

$$\text{Gain of the system} = A = \frac{e_0}{\Sigma e} = -\frac{\mu}{4 + \mu\beta}$$

For purposes of discussion of stability using the Nyquist criterion the above expression can be written in somewhat different way:

$$A = -\frac{\mu/4}{1 + \mu/4\ \beta}$$

This equation shows that the type of input circuit and parallel feedback we use has an effect of decreasing the gain by a factor of 4 as compared with a system using a single input and series voltage feedback when the equation for the gain is the widely known.

$$A = \frac{\mu}{1 + \mu\beta}$$

2. Output Impedance

The output impedance can be defined by the expression

$$Z_0 = \frac{e_0}{i_0}\bigg|_{\mathcal{E}_e = 0}$$

Under those conditions of zero input signal we are able to write the node equation at the output resistance.

$$\frac{E_0 - e_i}{R_4} + \frac{E_0}{R_2'} = i_0$$

Now we wish to express all voltages in terms of e_0. Using the fact that

$$E_0 \cong \beta e_0 \quad \text{and} \quad e_i = -\frac{e_0}{\mu}$$

it is possible to write

$$\frac{\beta e_0 + \frac{e_0}{\mu}}{R_4} + \frac{\beta e_0}{R_2'} = i_0$$

$$e_0\left(\frac{\beta + \frac{1}{\mu}}{R_4} + \frac{\beta}{R'_2}\right) = i_0$$

Then

$$Z_0 = \frac{e_0}{i_0} = \frac{1}{\frac{\beta + \frac{1}{\mu}}{R_4} + \frac{\beta}{R'_2}} = \frac{1}{\frac{\left(\beta + \frac{1}{\mu}\right)R'_2 + \beta R_4}{R_4 R'_2}} = \frac{R_4\ R'_2}{\left(\beta + \frac{1}{\mu}\right)R'_2 + \beta R_4}$$

Since $\mu \gg 1$ and with good approximation in most cases

$\beta \gg \frac{1}{\mu}$ We have

$$Z_0 = \frac{R_4 R_2'}{\beta(R_2' + R_4)} \cong \frac{R_2'}{\beta} \qquad \text{since } R_4 \gg R_2'$$

Where the resistance of the parallel combination R_2' and R_4 is practically R_2'. However

$$R_2' + \beta (R_1' + R_2')$$

Therefore $$Z_0 = \frac{\beta(R_1' + R_2')}{\beta} = R_1' + R_2'$$

which is actually the total load resistance placed at the output.

3. Input impedance

We will define the input impedance for the input 1 as

$$Z_1 = \frac{e_1}{i_1} \qquad \text{for } e_2 = e_3 = 0$$

Disregarding the source impedances as very small as compared with the input resistances it is possible to write:

$$i_1 = \frac{e_1 - e_i}{R}$$

As has been seen previously when the Gain was discussed

$$e_1 = \frac{\Sigma e + E0}{4} = \frac{\Sigma e + \beta e0}{4} = \frac{\Sigma e - \mu\beta ei}{4}$$

$$e_i = (1 + \frac{\mu\beta}{4}) = \Sigma e$$

$$e_i = \frac{\Sigma e}{\frac{4 + \mu\beta}{4}} = \frac{\Sigma e}{1 + \frac{\mu}{4}\beta}$$

For $e_2 = e_3 = 0$, we have

$$e_4 = \frac{e_1}{1 + \frac{\mu}{4}\beta}$$

And

$$i_1 = \frac{e_1 - \frac{e_1}{1 + \frac{\mu}{4}\beta}}{R_1} = \frac{e_1}{} \frac{\left(1 - \frac{1}{1 + \frac{\mu}{4}\beta}\right)}{R_1}$$

The input impedance will be given by

$$Z_1 = \frac{e_1}{i_1} = \frac{R_1}{1 - \frac{1}{1 + \frac{\mu}{4}\beta}}$$

This denominator is very nearly 1. Therefore, with good approximation, the input impedance for each input can be considered as given by the resistance placed in series with each input, namely

$$Z_1 \doteq R_1 \quad , \quad Z_2 \doteq R_2 \quad , \quad Z_3 \doteq R_3$$

C. <u>Contactor modulator and modulated waveform</u>

1. The contactor modulator

The contactor modulator, commonly known as "chopper", is, we may say, the starting point of our system.

It is essentially a single-pole, double-throw switch which vibrates at a frequency, called the carrier frequency, which is dictated by a voltage applied to its coil.

The amount of carrier voltage picked up by the signal circuit depends to some extent on the impedances used. For low impedance circuits

this amount is quite low.

In the medium impedance circuit it was used in, the pick up was appreciable and produced a zero error.

The constancy of duty ratio is assumed to be good in those devices. In our case, use of feedback and switch detection gave independence of duty ratio variations over reasonable limits.

The phase shift of the particular "chopper" we used, is about 110° at 400 cps, as said by the manufacturer. This phase shift is combined electrical and mechanical phase shift.

As can be seen from the circuit diagram only one side of the switch has been used to accomplish the modulation of the signal, which will be in consequence just half-wave modulated, giving from the start an attenuation of 2. Even if the type of input circuit used were suitable for full-wave modulation, it still would be worth not to use it, in face of the advantage of using the other side of the switch for the synchronizing circuit.

Except for the fact that the attenuation will increase the value of the minimum signal to be amplified, there is nothing else wrong with this loss since it can be compensated by the gain in the amplifier.

And the advantage of using the other side of the switch for synchronizing purposes is twofold:

a) Exact synchronization is possible at all times, independent of any phase disturbance at the "chopper" due to mechanical shocks or any other reason, since any disturbance in the signal carrier will affect the synchronizing voltage in exactly the same fashion.

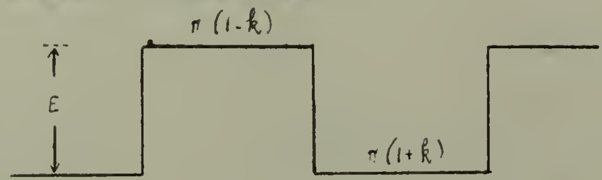
b) No power is drained from the carrier source for synchronizing purposes.

If the carrier were directly applied to the detector a definite amount of power would be required from the carrier source.

2. <u>Analysis of the input waveform</u>

An elementary Fourrier analysis will help to have a better understanding of the problem we have in hand when trying to amplify a switch modulated signal.

A wave of this form



Can be described by the following expression

$$f(t) = \frac{E}{2}(1-k) + \frac{2}{\pi} E \sum_{\mu=1}^{\infty} \frac{1}{2\mu-1} \operatorname{Cos}\left[k \frac{\pi}{2}(2m-1)\right] \sin\left[\omega_0 (2m-1)t\right]$$

$$- \frac{2}{\pi} E \sum_{n=1}^{\infty} \frac{1}{2n} \sin(k\pi n) \cos(2\omega_0 nt)$$

For $E = E_s \sin \omega t$ the expression will be modified:

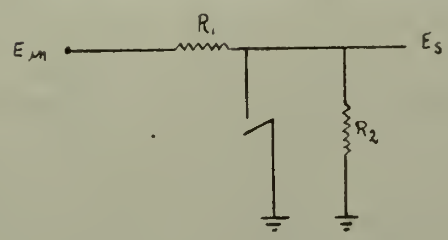
$$f(t) = \frac{Es \sin \omega t}{2} (1-k) +$$

$$+ \frac{2}{\pi} Es \sum_{m=1}^{\infty} \frac{1}{2m-1} \cos\left[k \frac{\pi}{2}(2m-1)\right] \frac{1}{2} \left[\cos\left\{\omega_0 (2m-1) - \omega\right\} t - \cos\left\{\omega_0 (2m-1) + \omega\right\} t\right]$$

$$- \frac{2}{\pi} Es \sum_{n=1}^{\infty} \frac{1}{2n} \sin(k\pi n) \frac{1}{2} \left[\sin\left\{\omega + 2\omega_0 n\right\} t + \sin\left\{\omega - 2\omega_0 n\right\} t\right]$$

This last expression shows that for the case when $k \neq 0$ we have the presence of every harmonic of the carrier frequency with the signal frequency added to or subtracted from it.

For the type of input circuit used



We have

$$E_s = \frac{R_2}{R_1 + R_2} E_{in}$$

When $R_1 = R_2$, $E_s = \frac{E_{in}}{2}$

The coupling capacitor found in our input circuit will prevent slow variations of grid current to reach the "chopper" and becoming modulated. Due to its value and to the time constant of the circuit, it will constitute practically a short circuit for 400 cps or higher frequencies. It will ofcourse stop d. c. and will not be an easy path for low frequencies.

The low capacitance condenser in parallel with the switch is designed essentially to absorb any self inductance effect coming from the switching action.

It will certainly attenuate high frequencies; this attenuation shows in the rounding of the leading edges of the otherwise square wave generated by the "chopper".

D. A. C. Amplifier and input circuit - Description of components

1. Introduction

The components of the signal circuit will be described from left to right as seen in the circuit diagram.

The values of the resistors and capacitors of a. c. amplifier will be commented upon only when they depart from the usual in some manner.

The choice of most of them was based on the tables found at the end of the RCA Receiving Tube Manual.

The sizes of the coupling and cathode capacitors were increased somewhat, the first one to decrease the phase shift at the carrier frequency and the second to lessen the possibility of "hum" pick-up across the cathode resistor.

2. Input resistances

The resistances used at the three different inputs and the feedback resistance were all chosen 1 M; the first three to insure the desired input impedance for the separate inputs and the fourth to make the feedback voltage to have the same weight as the input signals, in determining the input to the amplifier.

Cathode follower - The grid resistance of 1 M was chosen not to cause unnecessary attenuation to the input signal and its maximum value was limited by the desire of not increasing too much the impedance level at the input.

The value of the cathode resistor was chosen to give the desired value of bias at zero input signal. At the values used, the gain for small

signals is in the vicinity of 0.9.

3. Three stages of voltage amplification

The values of grid and plate resistances were chosen in accordance with the RCA Manual. Coupling and cathode capacitors were somewhat modified by the reason explained at the Introduction.

The values of cathode resistors were increased to bring down the operating points in the plate characteristics. The applied voltage Ebb came down in the initial stages due to the decoupling of this supply. In this way, using the same bias we would have smaller current and hence the larger value of the cathode resistors needed to secure that bias.

The current drain from the power supply was slightly decreased although this is not too significant. Linear operation was insured due to the small signal amplitude at the earlier stages. This linearity was verified later experimentally.

The size of the gain potentiometer was chosen as a regular grid resistor.

4. Output circuit

Under this title will be considered the 6L6 output tube and the output transformer.

Little can be told about the 6L6 since the instructions for use as class A_1 amplifier and 300 V plate supply were followed.

The output transformer was selected having ~~in~~ sight a good impedance matching, in order to make the 6L6 to operate on the desired load line. A transformer having the correct turns ratio (to match 15 K to 2500

ohms, i.e., 1:2.5, primary to half secondary) was not available. One having a larger turns ratio (1:3.5) had to be used and this made necessary an increase in the output impedance to maintain the ideal load at the 6L6 plate circuit.

5. Operating points

The operating points are indicated at the corresponding points in the circuit diagram.

E. Demodulator

1. Generalities

The ideal operation of the complete system we are trying to describe would be:

a) Generation of modulated square waves by a "chopper".

b) Amplification of the modulated square waves by a wide band amplifier.

c) Demodulation and exact reproduction of the original signal amplified by means of a second "chopper" in synchronism with the first one.

The power level we would have at the output precluded the use of a mechanical demodulator since usually these instruments are made for a current of the order of 1 ma. The field was wide open for the choice of a phase sensitive demodulator. Several types were investigated and finally the so called full-wave bi-directional switch demodulator was chosen.

This type of demodulator is the electronic equivalent to the mechanical switch and this fact primarily oriented the choice.

As can be seen from the diagram it has 8 diodes of which only 4 really matter in the question of linearity. Even so, use of a resistor in series with each diode acts as a compensation for difference in characteristics of the diodes in such way that the resistors themselves take the place of the vacuum tubes on the responsibility for balanced operation of the demodulator.

The circuit is very simple and good operation will not depend on the amplitude of the reference voltage as long as it is high enough to block the diodes in the presence of the applied modulated signal.

There is nothing to go wrong with the demodulator, except the diodes or a defective resistor.

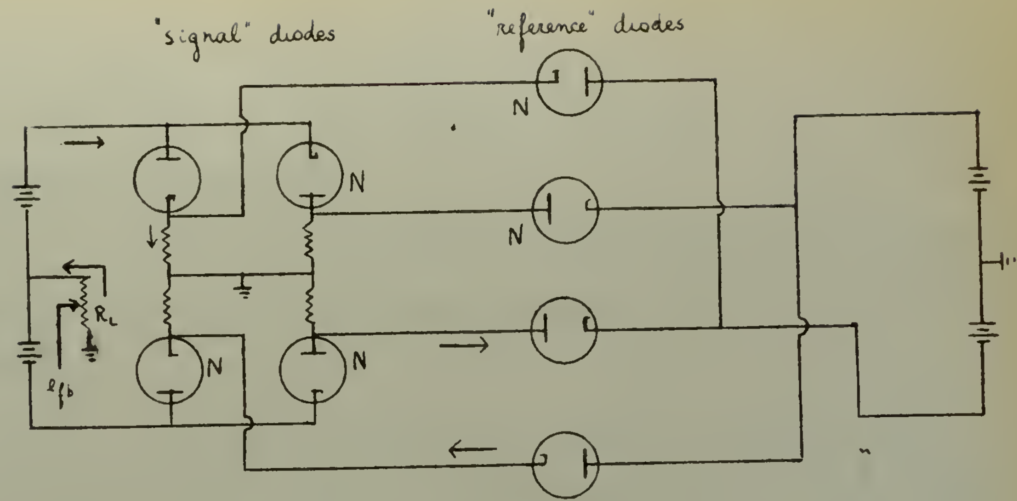
Once it is set in operation and the phase is adjusted for correct synchronism it will go on forever independent of further gain or balance adjustment.

The fact that it provides no gain does not constitute a drawback since the amplifier has plenty of reserve to supply it.

In systems where the ratio between the signal frequency and the carrier is relatively high this two way switch demodulator provides, beside the needed phase sensitivity, a rapid response, minimum carrier content and a reduced chance of interference of unwanted signals.

2. Operation

The following diagram will help the reader to have an idea of how the eight tubes are used to provide a full-wave phase-sensitive demodulation. A particular instant is chosen and at that instant the transformers are replaced by batteries. In order to understand the analogy with a switch action, it must be assumed that whenever the polarity of the reference signal changes, two of the "reference" diodes conduct instantly and the other two cut off also instantly. At the same time two of the "signal" diodes are blocked and two are unblocked also instantly. The reference voltage should have greater amplitude than the signal voltage in such way that the assumption can be considered as correct.



The letter N near a diode indicates "not conducting" either because there is no signal of the correct polarity applied or because it is blocked by the reference signal.

The diagram is self explanatory for every different situation and shows how only one half of the secondary of the signal output transformer works at a time.

3. Reflected loads

It is clear that under the ideal operating conditions, the output tube of the a. c. amplifier sees the total (internal and external) diode resistance of one diode in series with the load resistance reflected by one half of the secondary of the output transformer.

The simplifier diagram also shows that the output tube of the synchronizing circuit sees the total diode resistance of two diodes reflected by the whole secondary of the output transformer. It is easy to see

that the two transformers must differ if the two output tubes are the same and we want them to have identical operating conditions.

4. Maximum possible output

In our case we had to use identical transformers and this reduced the maximum obtainable reference voltage and consequently the maximum possible rectified signal.

It should be clear however that the physical limitation for the amplitude of the rectified signal is dictated by saturation of the diodes, provided the amplitude of the reference voltage is kept high enough to justify the assumption made previously. If a suitable combination of transformer and diode resistor were used we would be able to obtain a maximum output signal considerably larger than what was actually obtained (about 20 volts).

5. Data at medium signal level

In order to have an idea of the quality of operation of the demodulator in question, two 400 cps signals were applied at both ends: one of fixed amplitude at the reference end and one of controllable amplitude at the signal end. A 0.1 mf capacitor was used to smooth out the ripple.

Data and corresponding plot are contained in the next two pages.

It can be observed how nearly linear is the response at this voltage level. It can also be observed how the linearity falls off when the applied voltage approaches the reference voltage in magnitude.

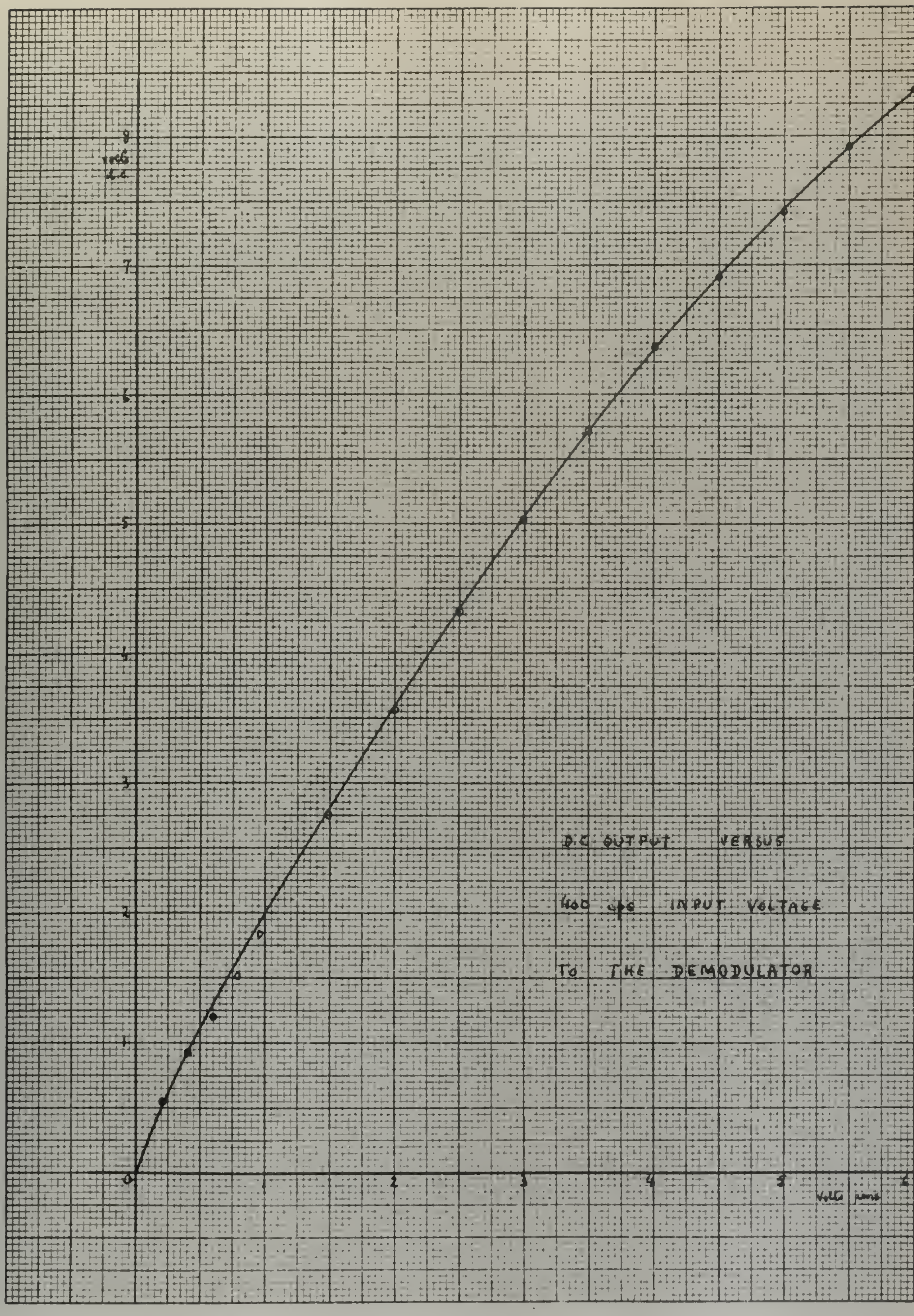
The performance of the demodulator in actual operation, integrated in the system, was good.

DEMODULATOR DATA

400 cps applied at both inputs

E ref. = 10.5 rms

E signal (volts rms)	E out (volts dc)	E ripple (volts rms)
.2	.55	.02
.4	.92	.022
.6	1.2	.024
.8	1.5	.03
1	1.77	.035
1.5	2.75	.05
2	3.55	.06
2.5	4.32	.075
3	5.02	.1
3.5	5.7	.12
4	6.37	.14
4.5	6.9	.16
5	7.4	.18
5.5	7.92	.21
6	8.35	.25
6.5	8.61	.28
7	8.8	.32



DC OUTPUT VERSUS
400 cps INPUT VOLTAGE
TO THE DEMODULATOR
volts d.c.
volts rms
0
1
2
3
4
5
6
7
8

F. Synchronizing circuit

1. Generalities

Following the same pattern as when the a. c. amplifier was described only the more unusual points will be commented upon.

Data for the stages of amplification using a 6SN7 and a 6L6 were taken with very little change from the RCA Manual.

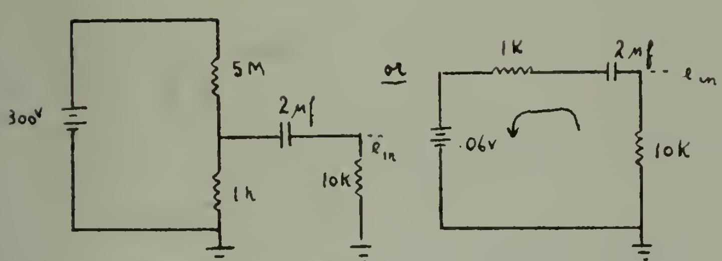
The load for the output tubes was not the correct one since a transformer with the desired turns ratio was not available, as has been pointed out in item E-4 of the section of Modulator.

2. Square wave generator

A circuit known generally as a sweep circuit was used at the input. Due to the large time constants involved as compared with the duration of the sweep, we really had a square wave generator. The duration of the sweep was controlled by the "chopper".

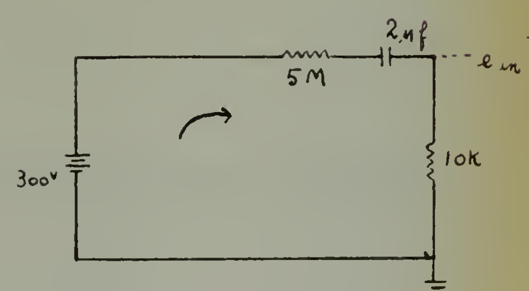
Depending on whether the switch is open or closed the circuit presents the following configurations:

switch closed



$T_{DIS} = 2 \times 11 \times 10^{-3} = 22 \times 10^{-3}$ sec.

switch open



$T_{CH} = 2 \times 5 = 10$ sec.

Where the period of charge and discharge are approximately the same and equal to 2.5×10^{-3} seconds.

Due to the relative magnitude of the time constants as compared with the periods of charge and discharge, a good approximation can be obtained by considering the current as constant instead of varying exponentially.

All we want is to have an idea of the magnitude of the input voltage to the amplifier.

Solution of two simultaneous equations involving exponentials give the value of the capacitor voltage at the beginning and end of charge and discharge, i. e., maximum and minimum capacitor voltages. They are respectively

$E_y = .76$ volts $\qquad\qquad E_x = .67$ volts

From the circuit diagrams for charge and discharge and using the assumption of constant current made previously, it is possible to calculate the two values of input voltage:

$$(e_{in})_1 = \frac{300 - .67}{5000 + 10} \times 10 \cong 0.6 \text{ volts}$$

$$(e_{in})_2 = \frac{.76 - .06}{10 + 1} \times 10 \cong -0.06 \text{ volts}$$

The 400 cps component must be amplified to be used as reference voltage at the demodulator.

3. Other significant features

Although the amplification factor of the 6SN7 is quite smaller than the one of the 6SL7, it still is more than enough for what we have

in mind. Assuming a gain of 150 for the two triodes and a gain of 15 for the output tube and output transformer the over-all gain is of the order of 2000 which would give a peak-to-peak amplitude of

$$(0.6 + 0.06) \times 2000 = 1320 \text{ volts}$$

This gain would be sufficient to saturate the output tube under those conditions.

For that reason a potentiometer was used as the grid resistor for the output tube, in such way that the gain may be adjusted for the best operation.

The cathode resistors for the two halves of the 6SN7 were somewhat different from the values presented in the RCA Manual because we were operating with a lower plate supply due to the decoupling network.

The capacitor used in paralled with the primary of the output transformer was experimentally determined to give the best synchronization for effective demodulation. It also serves the secondary purpose of attenuating high frequencies. In fact it also attenuates somewhat at 400 cps, but this is not serious since this can be more or less compensated by gain adjustment if the need arises for larger gain.

C. Power supply - Plate and Filament supply

1. Plate supply

Since no stabilized power supply was supposed to be used in this amplifier there was no particular need for building one.

At the start the 250 volts supply from the Laboratory was used. Since the amplifier was planned for 300 volts, a power supply existing in

the Laboratory and having the desired terminal voltage was used.

It presented good characteristics of regulation and filtering for the given load, and had no special features.

2. Filament supply

As dictated by previous experience, two types of heater supply were used: a. c. and d. c. Direct current supply was used for both 6SL7's while all the remaining tubes had 6.3 volts a. c. in their heaters.

The heaters of the 6SL7's were connected in series and a current of 0.3 amp. was made to flow across them from a bridge circuit of selenium rectifiers.

Effective filtering through electrolytic capacitors helped to reduce the ripple at the heater of the input tube to about 0.1 volt a. c. maximum. The good grounding of one side of this supply voltage had a considerable bearing in the performance of the system.

An insulating transformer was used to make this grounding possible.

No particular problem involved the 6.3 volts a. c. heater supply. It was made through a filament transformer. Both sides of the line were connected to ground across 50 ohms resistors.

FINAL REMARKS

A. Comments on the system

Use of modulation to improve the conditions of amplification of low frequency signals and use of feedback to make the gain less sensitive to changes in conditions of the circuit are supported by theory.

In fact the theoretical operation of the system being discussed is quite simple and all we have to worry about are the reasons which cause departure from the ideal conditions.

Despite the fact that the signal flow across the system is easily visualized, theoretical analysis is somewhat more involved due essentially to the presence of a non linear element, represented by the contactor. The contactor is said to be non linear, meaning that it can not be described by a Laplace Transform. However, if some analysis is to be made it should follow the same pattern presented on Reference 6.

Further difficulty on the analysis appears with the presence of a transformer whose characteristics are unknown and were not thought worthwhile to measure.

In that way all was tried to do was to limit the phase shift along the a. c. amplifier for the frequencies we were interested in, and to attenuate frequencies out of this range. This was accomplished through selection of adequate values for resistors and capacitors in the amplifier and in the output filter. It was assumed that the phase shift introduced by the output transformer wouldn't be large enough to become troublesome. Corrective networks as presented in page 339 of Reference 6 were not considered necessary.

B. Zero stability

The zero drift in the conventional direct coupled amplifier results from slow changes in the tubes.

Similar changes take place in the tubes in the a. c. amplifier also, but they are attenuated by the conventional R-C networks between tubes. The changes are further attenuated in the output transformer and finally due to the balanced arrangement of the demodulator circuit, they are reduced to a negligible value.

There are, however, other sources of zero disturbances and some of these are not so easy to eliminate. They can be of electrical, magnetic, mechanical or thermal nature. We will not be too much concerned with the last one since its level will be relatively low in comparison with the input voltages we intend to work with.

To improve the insensitivity to mechanical shocks, the first tube was mounted on a floating socket. If more improvement is desired the whole chassis can be mounted on rubber supports.

The electrical disturbances need some more attention. Since the signal goes through the amplifier essentially as a 400 cps. modulated voltage, any 400 cps. voltage entering the amplifier may disturb the zero. The safe practice would be to exclude the possibility of this happening. The only place it can enter the amplifier is through the contactor. At the start, the error coming from this fact was very noticeable but was decreased somewhat by the insertion of a center tapped 150 ohms at the output of the transformer supplying the 6.3 volts, 400 cps., to the contactor. The center

tap was connected to ground and adjusted to give the smallest zero error.

Because of the fact that the amplifier is to be used with up to 40 cps input signal, a 60 cps voltage in the amplifier will behave just like a signal and will be present in the output.

In an ideal set up, the input would have to be shielded and filtered to reduce the amount of 60 cps and 400 cps entering the amplifier.

Although a steady d-c current from the first grid cannot reach the "chopper", variations of it can. If they do they can become modulated and cause some interaction with the signal being modulated. Use of a cathode follower as the input tube improves the situation due to the very small changes in grid to cathode voltage in comparison with the corresponding changes of the input signal, from grid to ground.

Magnetic zero disturbances may be present due to the linking of loops in earlier stages of the amplifier with 60 or 400 cps magnetic fields. The 400 cps may cause a d-c error at the output and the 60 cps will be present as any other input signal. To correct this, the loops in the earlier stages must be as small as possible and if complete correction is desired, magnetic shielding would have to be used.

C. <u>Gain stability</u>

Elementary feedback theory shows that the changes which happen in a feedback circuit are decreased by a factor which is related to the loop gain. In this way, the larger the loop gain, the larger the gain stability we can achieve.

Stability problems prevented the system being discussed from having a loop gain larger than 40 at a forward over-all gain of 500. Under those conditions, making use of the expression $\frac{\mu}{1-\mu\beta}$ it is

easily seen that a change of 10% in μ will cause a change of only 0.3% on the net gain.

If the bandwidth of the system is reduced by the use of a large capacitor in the output, the loop gain may be increased. A 100μf capacitor brought the loop gain up to 50[illegible] but evidently in that case the bandwidth is reduced in a very great extent.

Under the conditions in which the amplifier was tested, it showed a good gain stability when was left for several hours with the same signal applied.

The response was found to be essentially flat from 0 to 40 cps.

D. <u>Stability against oscillation</u>

The maximum obtainable loop gain is limited by the possibility of oscillation and is a compromise with the bandwidth of the system. Use of a large output condenser will decrease the bandwidth but will help the stability and will permit use of a very large forward gain.

On the other hand if the bandwidth is to be increased the output condenser has to be decreased in size and a relatively large amount of 800 cps ripple is present in the output and this brings a definite limitation to the maximum loop gain. In other words a situation arrives when the 800 cps voltage is comparable in amplitude with the output voltage. When this happens the operation of the system becomes erratic.

Existence of feedback, other than the one intended, should be kept as low as possible. Every kind of spurious disturbance must be prevented. This was tried in several ways as decoupling of plate supply, filtering on the heater supply, good ground connection and shock proofing

of the input tube against mechanical feedback.

E. Performance of the system

No extensive measurement data were taken from the performance, since the system as it is is very flexible. Any amount of gain for frequencies from 0 to 40 cps can be obtained with varying values of gain stability, depending on the desired bandwidth.

Oscillation puts a definite limit on the optimum combination of gain stability and bandwidth. The enclosed data shows some of the results obtainable with the system.

On the data for verification of the condition

$$\left| \frac{\frac{e_0}{k} - \sum e_i}{\sum e_i} \right| \leq 0.02$$

Since the linearity and additive properties of the system had been verified previously, only the simplified expression below was computed.

$$\left| \frac{\frac{e_0}{k} - e_1}{e_1} \right| \leq 0.02$$

$$e_2 = e_3 = 0$$

Data was taken for $k = 10$ (minimum loop gain) since it constitutes the most critical situation. In other words, if the condition is satisfied for $k = 10$ it will also be for $k = 1, 2, 5$.

Three oscillographs for input signals of 20 cps, 30 cps, and 40 cps are enclosed. The net gain of the d. c. amplifier was 10 when those oscillographs were made. A very noticeable harmonic content can be seen in the 20 cps graph. 60 cps disturbance can be seen in all three.

The oscillographs were taken with a Brush 2-channel oscillograph.

DATA

Data for maximum and minimum net gain of the system when the forward gain was adjusted to give stable operation for a. c. signals under any condition.

20 cps — $e_{in} = 1$ V $e_{out} = 1$ V gain = 1

$e_{in} = .14$ V $e_{out} = 5.7$ V gain = 40.5

maximum loop gain = 40.5

30 cps — $e_{in} = 1$ V $e_{out} = 1$ V gain = 1

$e_{in} = .14$ V $e_{out} = 5$ V gain = 35.6

maximum loop gain = 35.6

40 cps — $e_{in} = 1$ V $e_{out} = 1$ V gain = 1

$e_{in} = .14$ V $e_{out} = 4.4$ V gain = 31.5

maximum loop gain = 31.5

D.C. — $e_{in} = 1$ V $e_{out} = -1.3$ V gain = 1.3

$e_{in} = .03$ V $e_{out} = -3$ V gain = 63.5

$e_{in} = 0$ $e_{out} = -1.1$V

equivalent input error = 0.0172 V

maximum loop gain = 49

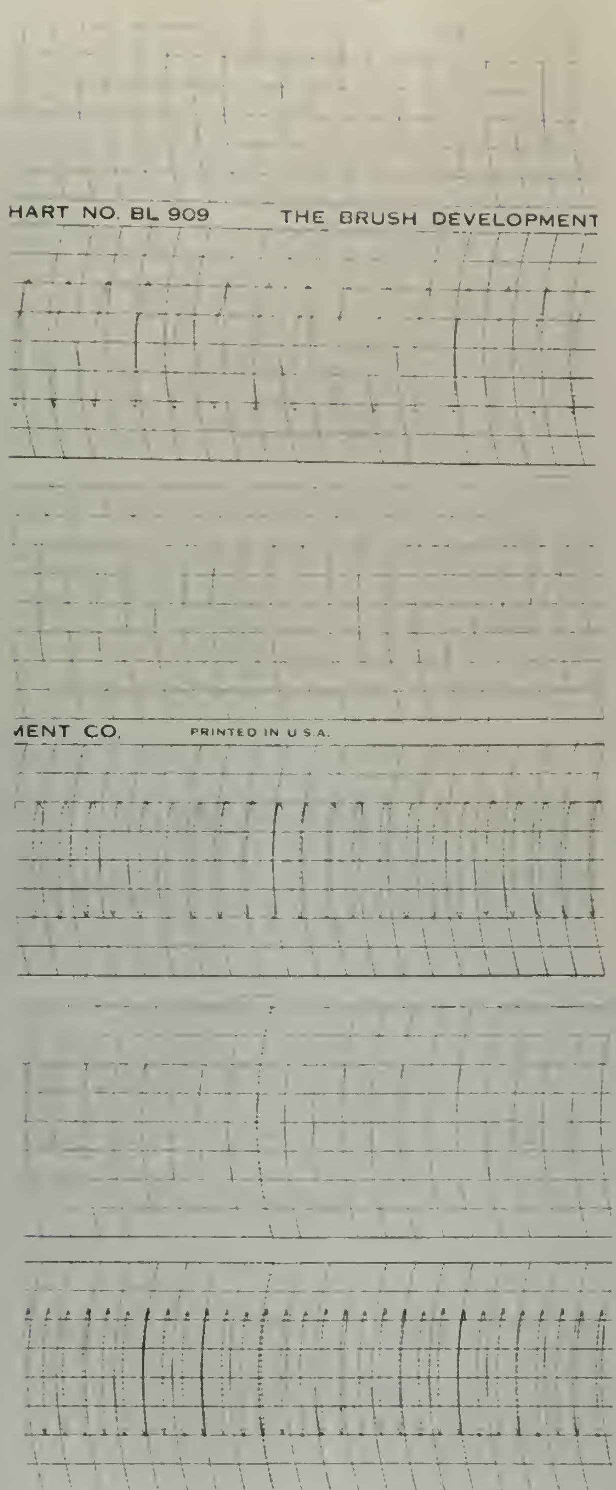
800 cps ripple at $e_{in} = 0$, open loop: $e_r = 0.09$V maximum

DATA

Data for computation of gain stability.

frequency	e_{in}(rms)	e_{out}(rms)	$\left\lvert \dfrac{\frac{e_0}{10} - e_{in}}{e_{in}} \right\rvert$
20cps	.2	2.05	.025
30cps	.2	2.05	0
40cps	.2	2	.025

Obs. The input voltage was accurately set on a Dumont 304-H Oscilloscope. The output voltage was read within .05 V rms from a RCA Junior Voltohmist VTVM.



20 cps

30 cps

40 cps

Oscillographs of input and output voltages of the d.c. amplifier with the net gain adjusted at the value of 10

CONCLUSIONS AND SUGGESTIONS

The amplifier as it was when the investigation was concluded can be considered as good. In fact it complied with all the specifications made at the start; some questions might still remain as for the flatness of gain, which was not extensively measured.

Definite limitation had to be placed on the forward gain due to stability reasons. The maximum possible loop gain was found to be of the order of 40. Small additions to the circuit and a more careful prevention of spurious signals should make possible the increase of the forward gain for the same net gain, what really means a better gain stability.

The zero d. c. error was decreased in some extent by a convenient ground on the 400 cps supply for the "chopper". Zero a. c. disturbances were curbed as much as possible.

A 60 cps voltage was present at the output. A better 60 cps electrical and magnetic insulation and shielding is necessary for future designs.

Introduction of a more elaborate filtering at the output will improve the system since the 800 cps ripple is a limiting factor for the minimum signal that can be applied to the input of the a. c. amplifier.

The "chopper" is considered to operate satisfactorily, the switch demodulator gave very good results and the synchronizing circuit proved to be a valuable addition since not once was the system observed to come out of synchronism.

Further work on this particular project must be centered on a more elaborate construction in order to prevent spurious voltages of coming

in and this will indirectly improve the system stability and decrease the zero disturbances.

Some modification made in the circuit in order to introduce a phase lead at the right frequency range, may allow for the use of a large forward gain and a flatter frequency response.

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